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Final Report — Volume One

# INVESTIGATION OF THE USE OF FREQUENCY-DIVISION MULTIPLE ACCESS FOR APPLICATION WITH A MIX OF USER TERMINALS

## Volume One — Summary and Results

By: PRAVIN C. JAIN

Prepared for:

DEFENSE COMMUNICATIONS AGENCY
SYSTEM ENGINEERING FACILITY
DEREY ENGINEERING BUILDING
RESTON, VIRGINIA 22070

CONTRACT DCA100-72-C-0033



STANFORD RESEARCH INSTITUTE
Menlo Park, California 94025 · U.S.A.

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I.

10 By: PRAVIN C./JAIN, V. Elaine/Hatfield, John K./Leung
D. Thomas/Magill

*Prepared for:*

DEFENSE COMMUNICATIONS AGENCY
SYSTEM ENGINEERING FACILITY
DEREY ENGINEERING BUILDING
RESTON, VIRGINIA 22070

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*Approved by:*

R. F. DALY, *Director*
*Telecommunications Department*

E. J. MOORE, *Executive Director*
*Engineering Systems Division*

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## ABSTRACT

This report presents the results of research performed to investigate the application of the frequency-division multiple-access (FDMA) technique for providing a mix of user terminals of differing characteristics--such as data rates, transmitter powers, and receiver sensitivities--with simultaneous access capability to a limiting satellite repeater. A computer program (SYSCON) has been developed to model a PSK/FDMA satellite communication system and to optimize its performance in operation with a mix of user terminals, through selection of power and frequency plans. This capability is achieved through optimization of a norm defined as the weighted sum of the link error rates and representing the figure of merit or a measure of the system's communications performance with respect to both the power and frequency of the links. The method of steepest descent is used for determining both power and frequency plans. It was found that through power and frequency control the limiting satellite repeater can be operated in the saturation region at substantially higher power levels (1-dB back-off) than is customary in practice.

To obtain the expression for the error rate at the input of the FDMA links, it was necessary to derive general analytic expressions for the limiter output signal, the intermodulation, and the noise components, when n signals are transmitted simultaneously through the satellite repeater. The expression for the bit error rate was then derived by assuming digital quadriphase modulation of the FDMA carriers and taking into consideration the presence of other FDMA carriers causing adjacent-channel interference, the intermodulation products generated in the limiter, and retransmitted satellite repeater noise, as well as receiver noise.

A general comparison of the three major multiple-access alternatives--FDMA, TDMA, and SSMA--with respect to selected performance criteria that are particularly important for the military environment and for operation with a mix of users indicated that FDMA performs much better than past analyses had shown. With the same satellite power and RF bandwidth, FDMA was found to offer nearly as much satellite throughput as TDMA and considerably more than SSMA.

CONTENTS

## ILLUSTRATIONS

TABLES

## ACKNOWLEDGMENTS

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The assistance of Professor M. E. Hellman of Stanford University and Dr. S. C. Fralick of SRI throughout this project was of major benefit.

# I INTRODUCTION

This report documents the results of a study performed to investigate the application of the phase-shift-keyed frequency-division multiple-access (PSK/FDMA) technique for providing simultaneous satellite multiple-access capability to a mix of user terminals with differing characteristics, such as data rates, transmitter powers, and receiver sensitivities. The exploration of PSK/FDMA potentials and capabilities was considered necessary because it was apparent that past analyses of FDMA had been overly pessimistic in regard to its capability as an efficient multiple-access alternative. The capabilities of this elementary and current-user-equipment-compatible technique seemed to be far from exhausted. It appeared that, through better understanding of its application with a limiting satellite repeater, its performance could be substantially improved, especially in conjunction with digital transmission techniques, which are expected to be the preferred mode of communication within the Defense Communications System (DCS).

## A. Computer Program SYSCON

A computer program (SYSCON) has been developed to model a PSK/FDMA communication system and to optimize its performance in operation with a mix of user terminals, through selection of power and frequency plans. The satellite transponder is modeled as a frequency-translating and limiting repeater whose limiting characteristic is described by a sum of two error functions. This analytic representation provides sufficient flexibility to describe accurately the measured transfer characteristics of practical limiters. The transmitters are characterized by their effective radiated power (ERP), RF frequency, and data rate. Quadriphase

PSK is assumed to be the data modulation for all users. The receivers are characterized by their antenna-gain-to-effective-noise-temperature ratio (G/T). The computer program is written in FORTRAN IV, conforming to the specifications of ANSI × 3.9--1966.

The capability to improve FDMA system performance is achieved through optimization of a norm representing the figure of merit or a measure of the system's communications performance. The optimization of the norm is done on either the frequency or the power at a time, while keeping the other fixed, and the procedure is then repeated on the other. The method of steepest descent is used in determining both frequency and power plans. At the end of the optimization, the best power and frequency assignments for the coincident users and the corresponding error rate for each FDMA channel are printed.

It is believed that this computer program will not only be a valuable tool in predicting and optimizing the performance of both existing and planned FDMA systems but that it can also be used very effectively to provide real-time control during actual system operation and for demand assignment, i.e., for allocation of the available communications channels to the various classes of users (fixed, mobile, small, or large) with differing priorities under dynamic conditions (communications demand changing with time).

From an operational standpoint, the combination of such a computer program with an efficient satellite monitoring technique will add an adaptive capability to the FDMA system for optimally matching the available satellite resources (power and bandwidth) to the user demands at all times, while ensuring that system performance degrades as little as possible, particularly when the available resources are reduced or degraded by such operational contingencies as jamming, environmental conditions, or partial system failure.

## B. Report Organization

The report consists of three volumes. Volume One provides a summary of the study which includes a description of the analysis approach, documentation of the pertinent equations, numerical results, and conclusions. Detailed analysis and investigation of the problem areas, as well as derivations of analytical expressions, are contained in the appendices in Volume Two. Three appendices represent essentially the results of Task 1, Task 2, and Tasks 3 and 4. Each appendix can be read independently. Thus it is not essential to read Appendix A to understand Appendices B and C, although this might be helpful. Volume Three provides a technical assessment of the suitability of PSK/FDMA for operation with a mix of users and compares its performance with other multiple-access alternatives, in particular, with that of time-division multiple access (TDMA) and spread-spectrum multiple access (SSMA).

## II CONCLUSIONS

The principal conclusion of this research effort is that the phase-shift-keyed frequency-division multiple-access (PSK/FDMA) technique has the potential to provide efficient satellite multiple-access capability to a mix of user terminals of differing characteristics and capabilities. The communication performance of an FDMA system can be improved substantially by providing power and frequency control to the transmitting terminals.

The major factors affecting the performance of an FDMA system are the losses resulting from power sharing and limiter suppression and the effects of cross products generated because of the mixing of signals in the limiter. Generally, power control of the transmitters will reduce the power-sharing and suppression losses, while frequency control will reduce the effects of the cross products.

For a given mix of user terminals, the best power and frequency plan can be determined from SYSCON, through optimization of a norm representing the figure of merit or a measure of the system's communication performance. The power and frequency plan thus obtained optimizes the norm for the selected ordering of the links in the repeater. Since some orderings are expected to be better than others, it is desirable to select a good ordering, and to determine the best power/frequency plan for that particular ordering. It has not been possible in this study to develop a theoretical approach for selecting the best ordering of the links in the repeater for a given mix of user terminals. However, a subroutine has been provided in SYSCON that allows consideration and optimization of all possible combinations of the links. Thus it is possible to determine the

best power and frequency plan for each of the possible link combinations, and then to select the best ordering.

A general comparison of the three major multiple-access alternatives, i.e., FDMA, TDMA, and SSMA, indicates that a controlled FDMA system can offer very nearly as much satellite throughput as TDMA and considerably more than SSMA. Since FDMA provides, in addition, simplicity, low cost, and compatibility with existing equipment, it could have substantial advantages over the other techniques.

TDMA and FDMA offer comparable performance for a mix of user G/T's and data rates if the same satellite bandwidth and power are available for both techniques. This result may appear surprising at first, since TDMA does not encounter the power-sharing and intermodulation effects that FDMA does. However, the result may be explained heuristically in the following fashion.

The mix in user G/T's causes the bandwidth utilization of a TDMA system to be reasonably low. Therefore, the required RF bandwidth is sufficiently large, and an FDMA system with the same throughput and bandwidth can avoid most of the cross products by selection of transmitter frequencies. Thus the FDMA system is not seriously degraded by this effect. In addition, a TDMA system may be expected to encounter some degradation from the presence of the local timing signals. Thus TDMA has a roughly compensating loss.

The power-control loss in an FDMA system might be estimated to be approximately 0.5 to 1.0 dB. No significant power-control loss occurs in most TDMA systems. However, in a TDMA system it is not possible to perfectly match burst rates with capacity quotients. A reasonable estimate of this loss is 0.7 dB. Because of its greater flexibility, an FDMA system can effectively avoid this loss. Consequently, with respect to throughput, TDMA and FDMA are roughly equivalent (to within a few decibels).

By contrast, SSMA is seriously hampered by its bandwidth inefficiency. Large differences in user G/T's and data rates force the required RF bandwidths to be excessive. Consequently, SSMA cannot be given serious consideration as a technique for high data rate transmission with a mix of user types. However, it must be noted that SSMA, as compared to TDMA and FDMA, has inherent AJ capability. Even with respect to this important performance criterion, however, SSMA has a significant failing. Without the use of a complex processing transponder, SSMA is degraded by the power-sharing effect in the satellite transponder. Thus SSMA is not nearly as effective as is theoretically possible. Nevertheless, it may provide the required order-wire and limited communications capability for "last-ditch" operation.

With FDMA (and SSMA) the primary problem is network power control, while with TDMA the significant problem is network timing. The difficulty with TDMA is that the control system is vulnerable to catastrophic failure. However, by careful initial system design, the occurrence of this can be made very improbable. By contrast, FDMA degrades gracefully, but significant degradation could occur much sooner with respect to small errors in network control.

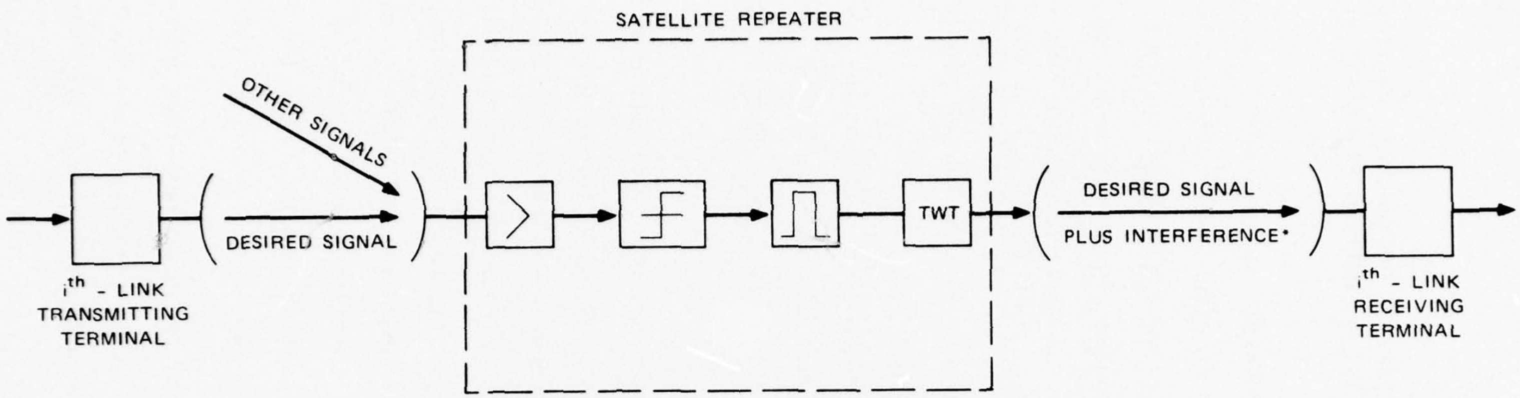
With respect to equipment, simplicity, and flexibility, FDMA is superior to TDMA. Consequently, it appears that the FDMA approach, since it yields technical performance comparable to that of TDMA, should be given serious consideration in military satellite communication.

## III PROBLEMS IN THE APPLICATION OF THE FDMA TECHNIQUE WITH A MIX OF USER TERMINALS AND A LIMITING SATELLITE REPEATER

In frequency-division multiple access (FDMA), signals from separate user terminals or links pass simultaneously through the satellite repeater and share the available repeater output power. Simultaneous occupancy by many signals in a common repeater is achieved by assigning different frequency bands to the several transmitting stations, so that they can be demodulated by frequency-selective receivers. Within its proper band, each multiple-access carrier can be modulated with any desired message modulation scheme, PSK, FSK, or FM. The major advantage of the FDMA technique lies in its simplicity; no network timing (as in TDMA) is required, and a channel can be selected by simple tuning of the receiver. The technique is fully compatible with existing user equipment hardware.

The general system configuration of an FDMA communication system is shown in Figure 1. The system comprises the transmitting and receiving terminals and the satellite repeater. The simplified version of the satellite repeater consists of a receiving antenna followed by a high-gain preamplifier, a limiter, a filter, a TWT amplifier, and, finally, the transmitting antenna. In practice, the repeater also contains a frequency translator, which, ideally, has no effect on the system performance. The frequency translator merely shifts the up-link frequencies of the signals by a specified amount for down-link transmission.

When FDMA is used with a nonlinear satellite repeater, its performance degrades because of the nonlinear characteristics of the limiter. The nonlinearity results in three potential sources of degradation.




*Interference consists of cross products, amplified signals of other users, and satellite repeater noise.

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FIGURE 1 GENERALIZED MODEL OF AN FDMA COMMUNICATION SYSTEM

## A. Power-Sharing Loss

Any nonlinear repeater, even the ideal automatic-gain-control (AGC) repeater, experiences a power-sharing loss effect. That is, the repeater output power is distributed to each access approximately (to within the suppression factor) in proportion to the ratio of that up-link signal power to the total up-link power. In principle, it is possible to set the up-link powers so that the desired output power is available for each access. This case would correspond to the case of zer[illegible]-sharing loss. In practice, this control cannot be accomplishe[illegible]fectly, and a power-sharing loss must be included in the link budget to account for the power imbalances.

## B. Suppression Loss

In operation with a mix of users, the number of signals and their power levels may vary constantly at the limiter input. This power imbalance can cause suppression of weak signals by strong ones, owing to limiting in the repeater. A strong signal having power greater than the sum of the powers of the other signals can capture the limiter, reducing the power output of the weak signals to anywhere from 1 to 6 dB less than that on a linear power-sharing basis, depending on the statistical amplitude distribution of the other signals.[1*] This undesirable consequence could be extremely critical for applications involving a mix of users with large and small transmitting and receiving capabilities.

The suppression of a weak signal is shown in Figure 2 for the case where the interference is a combination of a strong CW signal and Gaussian noise. When the limiter input consists of a large number of signals all

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*References are listed at the end of Volume One.

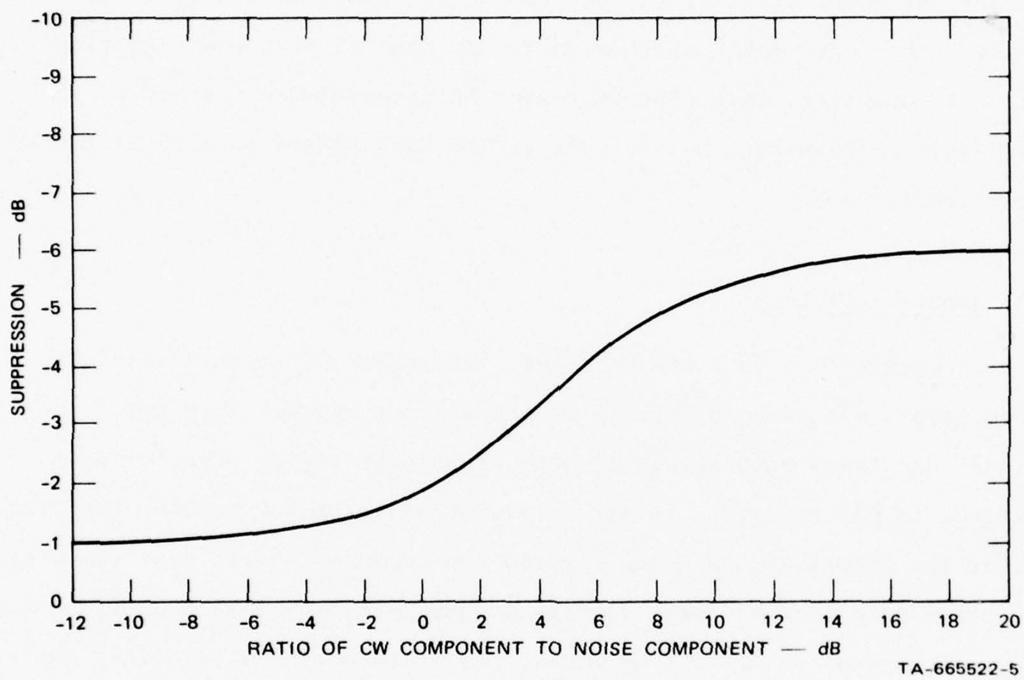


FIGURE 2 WEAK SIGNAL SUPPRESSION BY A COMBINATION OF SINUSOIDAL AND GAUSSIAN INTERFERENCE

approximately equal in amplitude except for one weak signal, the weak signal suppression will be 1 dB, since the amplitude distribution of all the other signals will approximate Gaussian noise. However, if one of the signals is very strong, it can suppress the weak signal by a maximum amount of 6 dB. It must be emphasized that such a large suppression is possible only if the power in the strong signal is much greater than the sum of the powers of all the other signals. In a practical system, this is usually feasible only when there are very few signals in the repeater, so one strong signal can capture the limiter. With the increase in the number of signals, the amplitude distribution at the limiter input would tend more toward Gaussian, and the suppression of the weak signal would drop to approximately 1 dB.

Power control of the transmitters will certainly alleviate the problem of power imbalance at the repeater and will also tend to preclude limiter capture by strong signals. From the above discussion, it is clear that the need for power control is occasioned not by the expectation of limiter capture, if many signals are usually present, but by the sheer power-sharing problem--exactly as though the repeater were linear. This makes the power-control problem substantially more tractable, since the need for precision in power control would be greatly alleviated with the increase in the number of signals in the repeater.

## C. Generation of Intermodulation Cross Products

Several signals simultaneously present in the repeater create intermodulation cross products. Only a small fraction of the repeater output power is wasted in these cross products. However, if some of the cross products fall in the frequency band of a desired signal, they interfere with the signal and add to the contribution from the up-link retransmitted satellite noise and down-link receiver noise.

The problems of signal suppression and the generation of cross products are, of course, related, since both are direct consequences of limiting in the satellite repeater. The combined effect of weak signal suppression and the presence of intermodulation components on the down link will reduce the effective signal-to-noise ratio (SNR) at the receiving terminal. If the SNR falls below the threshold required to maintain a specified error rate, the communication link will be unusable.

It is well known[2] that when the frequency assignments of multiple-access carriers are spaced closely and uniformly across the available repeater bandwidth, so that the amplitude distribution of the composite signal closely approximates Gaussian noise at the repeater input, the signal-to-intermodulation-power (S/I) ratio in the central channel is approximately 9 dB, while for the channels at the edge of the band it is typically 1 to 2 dB higher. A value of only 9 dB for S/I is certainly low for analog modulation, since it is in the vicinity of threshold for such systems as wideband FM. In addition, the presence of up-link retransmitted noise and down-link receiver noise will lower the effective SNR at the receiver even more. The problem is further accentuated by the fact that in operation with a mix of users, because of limiter suppression of weak signals, the S/I for a weak signal could easily fall below an acceptable value.

A 9-dB S/I, however, can be quite acceptable for digital transmission, since it is sufficient to achieve an error rate of approximately $10^{-5}$ with a binary phase-shift-keyed (BPSK) modulation scheme. The S/I for all the other channels will be greater than 9 dB, which will yield error rates smaller than $10^{-5}$. The S/I can be improved through use of nonuniform frequency plans, so that a larger fraction of the intermodulation products falls on unoccupied channels. For example, by using only 50 percent of available bandwidth, which is equivalent to reducing the

number of communication channels by two, it would be possible to achieve an improvement in S/I of at least 3 dB. The loss in the number of channels could be made up by employing a more efficient modulation scheme, such as QPSK (quadriphase) rather than BPSK. Thus, with a QPSK/FDMA system, it would be possible to obtain the same number of channels in the available repeater bandwidth, but with an S/I of 12 dB rather than the 9 dB that would be obtained with a BPSK/FDMA system. This increase in S/I can be used as additional margin for combatting retransmitted up-link as well as down-link noise and other interference. This margin could be further increased by providing power control.

## D. Summary

Power-sharing, suppression, and intermodulation cross products present difficulties for FDMA operation with a nonlinear satellite repeater. However, these impairments are by no means characteristic of FDMA alone. In fact, any multiple-access technique that involves simultaneous presence of user signals in a limiting repeater will suffer from these drawbacks; this is true for both FDMA and SSMA. Network power control can significantly reduce the magnitude of the first two degradations. The effect of cross products can be reduced by frequency control. Nominally, the anticipated intermodulation cross-product power levels for digital modulation systems are tolerable.

# IV THEORETICAL INVESTIGATION OF A PSK/FDMA SYSTEM WITH A MIX OF USER TERMINALS

From the discussion in Section III, it is evident that an investigation of the problems involved in the application of FDMA with a limiting satellite repeater would be of considerable practical interest and importance in the design and operation of the communication system. It is reasonable to expect that, for any given mix of user terminals, there exists an optimum power and frequency plan that would provide the best utilization of the satellite resources consistent with obtaining the best system performance. While this concept is widely acknowledged, it has not been possible in the past to find such a power and frequency plan.

This section is intended to summarize the approach and results of our investigation of the effects of limiting in the satellite repeater on the performance of a QPSK/FDMA system in operation with a mix of terminals, and to describe our approach for selecting frequency and power assignments for the transmitting terminals, so as to achieve the best utilization of the FDMA system capabilities. The investigation was confined to analysis of the problems resulting from the nonlinear characteristic of the limiter only.* Other sources of system degradation, such as intersymbol interference caused by filtering and AM-to-PM conversion introduced by the TWT amplifier, were not considered.

We begin with a description of the model of the satellite repeater.

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*Detailed analysis and derivation of equations are contained in the appendices in Volume Two.

## A. Model of the Satellite Repeater

A bandpass limiter (see Figure 3) is generally used in theoretical investigations to model a limiting satellite repeater. The limiting can be either hard or soft. The bandpass filter preceding the limiter is wide enough to pass the signals with negligible distortion; it limits the repeater input noise to a bandwidth that is small compared to the center frequency of the filter. The bandpass filter following the limiter

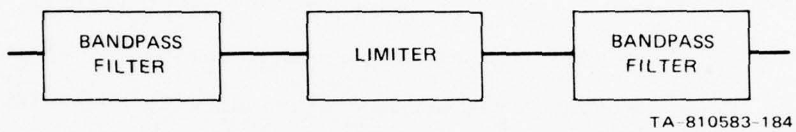


FIGURE 3 SATELLITE REPEATER MODEL FOR INVESTIGATION OF IDEAL SYMMETRIC LIMITING

confines the output spectrum to essentially only the fundamental band of the signals. The filter preceding the limiter has the same bandwidth as the high-gain preamplifier in Figure 1. Thus the SNR at the limiter input in Figures 1 and 3 will be the same. The gain of the limiter in Figure 3 is identical to that of the combination of the limiter and TWT in Figure 1. It is assumed that both the preamplifier and the TWT amplifier in Figure 1 are linear, so all the nonlinearity existing in the satellite repeater can be attributed to the presence of the limiter alone.

The model of Figure 3 requires an "ideal" zonal filter at the limiter output, to pass the fundamental band of signals unattenuated but to completely suppress the higher harmonics of the signal frequencies. Real filters can only approximate this desired behavior.

B. Limiter Characteristics

To analyze the problem of transmitting a number of constant-envelope, phase-modulated sinusoidal carriers through the bandpass limiter shown in Figure 3, a mathematical representation is needed for the nonlinear characteristics of the limiter.

In practice, a limiter is generally characterized by its power transfer characteristic. Figure 4 shows the measured power input/output transfer characteristics of a tunnel diode limiter.[3] Both the input and the output power levels are normalized to a maximum limiter power output equal to 0 dBW. The dashed line in Figure 4 shows the normalized power transfer characteristic of a linear amplifier of unity gain. At low

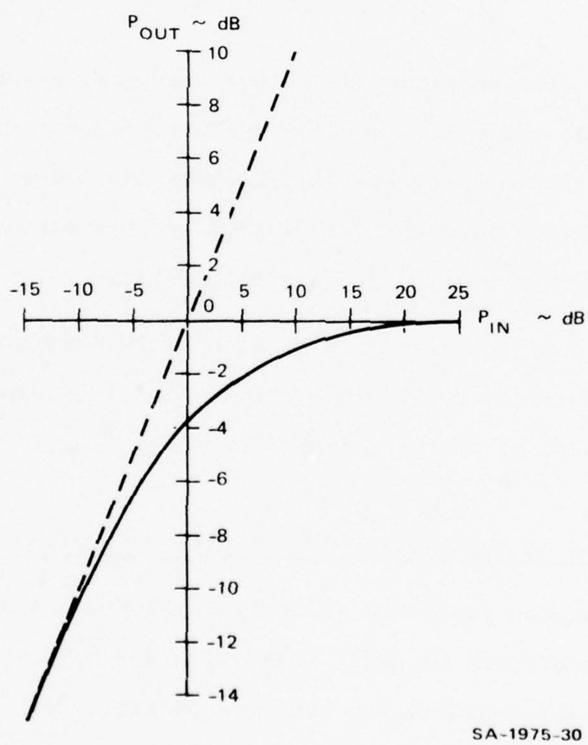


FIGURE 4 MEASURED POWER TRANSFER CHARACTERISTIC OF A TUNNEL-DIODE LIMITER

input power levels, the limiter output power level increases linearly (unity slope) with the input. As the input power level is increased, the output departs from the linear characteristic and reaches saturation at large input power levels.

To represent the transfer characteristics of practical limiters, such as shown in Figure 4, the analytical representation must provide unity slope at low signal power levels, with an additional parameter available to characterize the softness with which the limiter reaches saturation. This flexibility can be achieved by describing the limiter voltage, characteristic by a sum of two error functions (see Appendix A, Section 2-d):

$$y = \frac{\alpha}{2}\left[\mathrm{erf}\left(\frac{x}{\sqrt{2}\gamma_1}\right) + \mathrm{erf}\left(\frac{x}{\sqrt{2}\gamma_2}\right)\right] \quad , \qquad (1)$$

where x and y are limiter input and output voltages, respectively; $\alpha$ is the limiting level; and erf( ) denotes the error function. The parameters $\gamma_1$ and $\gamma_2$ can be chosen appropriately, so that the power transfer characteristic as determined from Eq. (1) closely approximates the measured power transfer characteristic of practical limiters (such as in Figure 4).

We consider next the transmission of n angle-modulated FDMA carriers through a bandpass limiter (Figure 3) whose limiting characteristic is described by the sum of two error functions [Eq. (1)].

## C. Determination of the Limiter Output with n Signals and Noise

Stanford Research Institute recently developed a time-domain method of analysis[4] that differs markedly from the older mathematical techniques that have been widely used in the analysis of limiters.[5-8] The main feature of this analytical approach is that it permits the limiter output--consisting of signals, intermodulation products, and noise--to be expressed

completely in the time domain. Representation in the time domain permits retention of the phase of the FDMA signals and intermodulation products at the limiter output. This is not possible, in general, with frequency-domain techniques and, in particular, with the autocorrelation method, which is usually employed in limiter studies. The time-domain technique also permits evaluation of the distribution of the interfering signals (cross products and noise) at the limiter output, which is not possible with the autocorrelation function approach.

1. Amplitudes of the Output Signal and Intermodulation Products

Appendix A (Section 3) shows that, if the limiter characteristic is represented by Eq. (1) and if its input consists of n sinusoidal carriers of arbitrary amplitude and phase modulation,

$$s(t) = \sum_{i=1}^{n} a_i \cos\left[\omega_i t + \phi_i(t)\right] \quad , \tag{2}$$

plus a narrow band of stationary Gaussian noise, the limiter output in Figure 3 prior to bandpass filtering can be expressed as

$$z(t) = \frac{\alpha}{2} \sum_{p_1=-\infty}^{\infty} \cdots \sum_{p_2=-\infty}^{\infty} h_{p_1 p_2 \ldots p_n} \sin\left\{\sum_{i=1}^{n} p_i \left[\omega_i t + \phi_i(t) + \frac{\pi}{2}\right]\right\} \quad , \tag{3}$$

where

$$h_{p_1 p_2 \ldots p_n} = \frac{2}{\pi} \int_0^{\infty} \prod_{i=1}^{n} J_{p_i}\left(v a_i\right) \left\{\exp\left[-\frac{v^2\left(\sigma^2 + \gamma_1^2\right)}{2}\right] + \exp\left[-\frac{v^2\left(\sigma^2 + \gamma_2^2\right)}{2}\right]\right\} \frac{dv}{v} \tag{4}$$

represents the amplitude coefficient of the output signal and intermodulation components. The significance of the parameters in Eq. (4) is as follows:

- n is the total number of signals (links).
- The value of $p_i$ is the harmonic of the $i^{th}$ signal appearing in the cross product under consideration. Thus, $p_i = 0$ indicates that the $i^{th}$ signal does not contribute to this cross product, while $p_i = 1$ indicates that the fundamental appears, and so forth. The composition of the cross product is indicated by the subscript $p_1 p_2 \ldots p_n$. A subscript 100...0 indicates that the output term under consideration is Signal 1. A subscript 2100...0 indicates that the second harmonic of Signal 1 is mixed with the fundamental of Signal 2, and so forth.
- $a_i$ is the amplitude (voltage) of the $i^{th}$ input signal.
- $\sigma^2$ is the noise power at the input to the limiter. Thus, $a_i^2/2\sigma^2$ is the SNR of the $i^{th}$ link at the input to the limiter.
- $\omega_i$ is the radian frequency of the $i^{th}$ signal.
- $J_{p_i}(va_i)$ is the Bessel function.
- $\alpha$ is the limiting level in saturation.
- $\gamma_1$, $\gamma_2$ are the parameters in Eq. (1).

2. <u>Frequencies of the Intermodulation Products</u>

Equation (3) shows that the limiter output contains components whose frequency is given by:

$$f = p_1 f_1 + p_2 f_2 + \ldots + p_n f_n \quad , \tag{5}$$

where

$$p_1, p_2, \ldots, p_n = (0, \pm 1, \pm 2, \ldots).$$

Shaft[4] has shown that the dominant cross products are those for which $|p_i|_{max}$ is small ($\leq 2$). This approximation considerably simplifies the determination of the frequencies, since, instead of considering all the $p_i$ from $-\infty$ to $\infty$, it is necessary to consider them only in the range from -2 to 2.

We are interested here only in the determination of the cross products that fall in the pass band of the filter following the limiter, i.e., those falling in the first zone,

$$p_1 + p_2 + \ldots + p_n = 1 \quad . \tag{6}$$

All the other cross products will be filtered out by the bandpass filter.

Since the limiter has an odd-order transfer characteristic, generally only the third-order intermodulation components are the most significant contributors to the output distortion.[4] For the third-order intermodulation products, the sum of the magnitudes of the $p_i$ coefficients is three,

$$|p_1| + |p_2| + \ldots + |p_n| = 3 \quad . \tag{7}$$

With the aid of Eqs. (6) and (7) we can determine all the third-order cross products that will pass through the output bandpass filter (Figure 3). Note that there are two types of third-order cross products. These are generated at frequencies

$$2f_i - f_j = f_{21} \quad , \qquad |p_i| = 2,\ |p_j| = 1$$

and

$$f_i + f_j - f_k = f_{111} \quad , \qquad |p_i| = |p_j| = |p_k| = 1 \quad .$$

The cross product at the frequency $2f_i - f_j$ is produced by the mixing of two signals at frequencies $f_i$ and $f_j$, respectively. On the other hand, the cross product at the frequency $f_i + f_j - f_k$ is generated by the mixing of three signals at frequencies $f_i$, $f_j$, and $f_k$.

3. Limiter Output Noise Components

In FDMA, the total power in all the signals is much larger than the power in the thermal noise at the limiter input. Appendix A (Section 3) shows that the distribution of the noise at the limiter output then tends to a Gaussian distribution:

$$p(n) = \frac{e^{-n^2/2\sigma_n^2}}{\sqrt{2\pi\sigma_n^2}} \quad , \tag{8}$$

whose variance is given by:

$$\sigma_n^2 = \frac{2}{\pi}\left[\sqrt{\frac{\alpha^2}{2\left(\sigma_2^2 + \gamma_1^2\right)}} + \sqrt{\frac{\alpha^2}{2\left(\sigma_2^2 + \gamma_2^2\right)}}\right]^2 \cdot \sigma^2 \quad , \tag{9}$$

where $\sigma^2$ is the total noise power at the limiter input, and $\sigma_2^2$ represents the total power of all the input signals, i.e.,

$$\sigma_2^2 = \sum_{i=1}^{n} a_i^2/2 \quad .$$

We have thus obtained all the analytic expressions that are needed to determine the limiter-output-signal, cross-product, and noise components when a sum of n signals is transmitted simultaneously through the repeater. Next, we consider the problem of demodulation of the FDMA signals at the receivers after transmission and amplification through the satellite repeater.

## D. Error Rate at the Output of an FDMA Channel

It is assumed that all the transmitters in the system use quadriphase modulation of the FDMA carriers and that the signals at the receivers are detected by a correlation operation with a synchronized reference.

At the receiver, the desired signal is detected in the presence of (1) other FDMA signals causing adjacent channel interference, (2) cross products generated as a result of the mixing of the signals in the limiting repeater, and (3) retransmitted up-link (satellite-repeater) as well as down-link (receiver front-end) noise. Appendix B shows that the error rate at the output of the $i^{th}$ channel is given by:

$$P_{ei} = \frac{1}{2}\left[1 - \mathrm{erf}\left(r_i/\sqrt{2}\right)\right] \quad , \tag{10}$$

where erf( ) is the error function, and

$$r_i^2 = \frac{\rho_i^2}{1 + \sum_{\substack{k=1 \\ k \neq i}}^{n} \rho_{ik}^2 \, S_{ik} + \sum_{j=1}^{m} \hat{\rho}_{ij}^2 \, \hat{S}_{ij}} \tag{11}$$

represents the equivalent SNR or $2E/n_o$ at the output of the $i^{th}$ channel, and where

$n$ = number of limiter input signals (links)

$m$ = number of cross products falling in the repeater bandwidth

$\rho_i^2$ = SNR of the desired signal at the $i^{th}$ receiver

$\rho_{ik}^2$ = SNR of the $k^{th}$ signal at the $i^{th}$ receiver

$\hat{\rho}_{ij}^2$ = intermodulation-to-noise ratio of the $j^{th}$ cross product at the $i^{th}$ receiver.

The parameters $S_{ik}$ and $\hat{S}_{ij}$ represent the effect of integrate and dump filtering in the $i^{th}$ receiver on the $k^{th}$ interfering link and on the $j^{th}$ cross product, respectively.

The SNRs $\rho_i^2$, $\rho_{ik}^2$, and $\hat{\rho}_{ij}^2$ can be expressed in terms of satellite and receiver parameters as (Appendix C, Section 5):

$$\rho_i^2 = \frac{2P_r \cdot \left(G_{ri}/T_{ri}\right)}{K \cdot R_i} A_i^2 \tag{12}$$

$$\rho_{ik}^2 = \frac{2P_r \cdot \left(G_{ri}/T_{ri}\right)}{K \cdot R_i} A_k^2 \tag{13}$$

$$\hat{\rho}_{ij}^2 = \frac{2P_r \cdot \left(G_{ri}/T_{ri}\right)}{K \cdot R_i} A_{m,n,p}^2 \quad , \tag{14}$$

where

$P_r$ = satellite ERP referenced to the ground terminal

$G_{ri}/T_{ri}$ = figure of merit of the $i^{th}$ receiver

$K$ = Boltzman's constant

$R_i$ = bit rate of the $i^{th}$ link

$A_{p_1 p_2 \cdots p_n} = \pi/4 \cdot h_{p_1 p_2 \cdots p_n}$.

The A parameters are identical with the h parameters of Eq. (4), except for the factor $\pi/4$. In Eqs. (12) through (14) the indices i, k, m, n, and p indicate the links involved, while j is an index on the intermodulation products. Thus, $A_i$ and $A_k$ denote the amplitudes of the signals for the $i^{th}$ and $k^{th}$ links, respectively, while $A_{m,n,p}$ represents the amplitude of the third-order cross product generated by the mixing of the signals on links m, n, and p.

The parameters $S_{ik}$ and $\hat{S}_{ij}$ may be determined from the following expressions:

$$S_{ik} = \int_{-\infty}^{\infty} |H_i(f)|^2 \cdot G_k(f)\, df \tag{15}$$

$$\hat{S}_{ij} = \int_{-\infty}^{\infty} |H_i(f)|^2 \cdot \hat{G}_j(f)\, df \quad , \tag{16}$$

where

$H_i(f)$ = transfer function of the $i^{th}$ integrate and dump filter.

$G_k(f)$ = power spectral density of the $k^{th}$ signal (normalized by the power of the $k^{th}$ signal).

$\hat{G}_j(f)$ = power spectral density of the $j^{th}$ cross product (normalized by the power of the $j^{th}$ cross product).

Alternatively, by using Parseval's theorem, we can determine $S_{ik}$ and $\hat{S}_{ij}$ in terms of autocorrelation functions as

$$S_{ik} = \int_{-\infty}^{\infty} R_{h_i}(\tau) \cdot R_{\Phi_k}(\tau)\, d\tau \tag{17}$$

$$\hat{S}_{ij} = \int_{-\infty}^{\infty} R_{h_i}(\tau) \cdot R_{\psi_j}(\tau)\, d\tau \quad , \tag{18}$$

where the autocorrelation functions are defined as

$$R_{h_i}(\tau) = \text{Fourier transform of } |H_i(f)|^2$$

$$R_{\phi_k}(\tau) = \text{Fourier transform of } G_k(f)$$

$$R_{\psi_j}(\tau) = \text{Fourier transform of } \hat{G}_k(f).$$

The evaluation of $S_{ik}$ and $\hat{S}_{ij}$ is discussed in Appendix B.

Since by Eq. (10) the probability of error is a monotonic function of the SNR, any improvement in the receiver output SNR will also improve the channel error rate. The dependence of the probability of error on the frequency plan is displayed in the two integrals involved in the determination of $S_{ik}$ and $\hat{S}_{ij}$, while the dependence on the power plan is displayed in the SNRs, $\rho_i^2$, $\rho_{ik}^2$, and $\hat{\rho}_{ij}^2$, since these are functions of the A parameters, which, in turn, are determined by the ERPs of the transmitters.

Next, we consider an approach for assigning powers and frequencies to the transmitters for improving the performance of the FDMA system.

## E. Approach for Selection of Power and Frequency Plans

When several FDMA links are simultaneously active, they must share the limited resources of the satellite. These resources are described in terms of the satellite effective radiated power (ERP) and the repeater RF bandwidth. It is desirable that the available satellite power and bandwidth be shared optimally among the coincident users and that the communication efficiency of the FDMA system be maximized.

As mentioned in Section III, when several links occupy the repeater simultaneously, the limiting in the repeater introduces cross products which may interfere with the reception of the signals. The impact of an intermodulation product depends upon its frequency location and its

amplitude relative to the desired signals. The effect of the cross products can be varied by controlling the frequency and power of the transmitted signals. Decreasing power causes the limiter to saturate less and thus reduces the amplitude of the cross products relative to the signal power; however, it also reduces the absolute power in the signals and hence reduces the SNR. Thus the selection of signal power is determined by the tradeoff between the effects of noise and the effects of cross products.

Changing the signal frequencies changes the frequencies of the cross products. Thus, it is possible to choose carrier frequencies that reduce the effect of intermodulation products on a particular signal by increasing the distance (in frequency) of the cross products from the signal carrier. However, because of the limited repeater bandwidth and the fact that many signals are present, any change in frequency may help one signal at the expense of another.

The problem, therefore, is to develop an approach for selecting frequency and power plans that will provide, for a given mix of user population, the best utilization of the FDMA system capability. The search for good power and frequency plans should also consider the constraints imposed by the characteristics of the repeater and the terminals. These may be specified in terms of:

- Limiter characteristic
- Satellite repeater bandwidth
- Satellite noise temperature
- Satellite ERP
- Data rates between pair of ground terminals
- Number of links

- Receiver figure of merit (G/T)
- ERP of the ground terminals.

Mathematically, the problem for determining power and frequency plans can be formally stated as follows: Given the satellite and terminal characteristics, find a power and frequency plan such that a selected norm that represents the figure of merit or performance of the FDMA system is optimized.

### 1. Definition of Norms

To talk quantitatively about optimization, it is essential to define a measure of system performance or norm for the system. Three norms suggest themselves; each is consistent with a different system philosophy:

- *A simple norm can be defined as the sum of the* link error probabilities:

$$N_a = \sum_{i=1}^{n} P_{ei} \quad . \tag{19}$$

  This norm would treat all signals equally and would not account for the fact that some carry information at a higher rate than others.

- To account for the fact that some signals carry more information than others, a norm can be defined as the sum of the error probabilities weighted by the data rates:

$$N_b = \frac{\sum_{i=1}^{n} P_{ei} \cdot R_i}{\sum_{i=1}^{n} R_i} \quad . \tag{20}$$

Minimization of this norm would minimize the average error rate and maximize the information transfer rate.

- Finally, one can take the minimax approach to ensure that the poorest channel has as good performance as possible. This leads to the norm:

$$N_c = P_{ei\ max} \quad . \tag{21}$$

Minimizing this norm will minimize the maximum error probability and ensure that every other channel meets or exceeds this performance.

## 2. Description of the Selected Approach

The approach taken for determining the best power and frequency plans, for a given mix of user terminals, is to optimize a selected norm first with respect to either the frequency or the power, holding one of them fixed, and then to repeat the procedure on the parameter that was held fixed. The reason for doing this is that frequency and power are easily separable in the expression of the error rate in Eq. (8). The dependence of the probability of error on the frequency plan is displayed in the two integrals involved in the determination of the parameters $S_{ik}$ and $\hat{S}_{ij}$, while the dependence on the power plan is displayed in the SNRs, $\rho_i^2$, $\rho_{ik}^2$, and $\beta_{ij}^2$, since these are functions of the transmitter ERPs.

The problem of minimizing the norm with respect to power or frequency is equivalent to finding the minimum of a function of n

variables. A well-proven technique for finding the local minimums of a multidimensional surface is the method of steepest descent. It consists of picking a point, calculating the partial derivatives at this point, and proceeding in the direction of the steepest slope by a specified amount. After the step has been taken, the function is evaluated to see whether it is less than its previous value. If it is less, the procedure is continued; if not, a smaller step is taken. A local minimum will eventually be found if it exists.

It should be mentioned that, when a starting point has been selected, this approach merely ensures convergence to the minimum closest to the starting point. If the surface contains many minimums, the selection of the initial starting point is very important, because this determines whether we converge to a local minimum or the global minimum (the minimum of all the minimums).

The approach used in both power and frequency search is an improved version of the method of steepest descent described in the following. More detailed description is provided in Appendix C, Section 6.

In the following discussion, $N(\tilde{\Phi})$ is designated as the objective function corresponding to the norm to be optimized, and $\tilde{\Phi}$ is either the power or the frequency of the input signals. An initial value of the selected norm is calculated at the start of the optimization procedure by specifying the powers and the frequencies of the transmitters. Next, the partial derivatives of the norm $\partial N(\tilde{\Phi})/\partial\Phi_i$ are calculated at these values of $\tilde{\Phi}$. The calculation of the derivatives of the norm with respect to the signal amplitudes and frequencies is accomplished by using the analytic expressions for the partial derivatives.

The new values of $\tilde{\Phi}$ are then determined in any iteration j as

$$\tilde{\Phi}^{j+1} = \tilde{\Phi}^{j} + \tilde{\alpha}^{j}\tilde{S}^{j} \quad , \qquad j = 0,1,2,3,\ldots,m \quad , \qquad (22)$$

where $\tilde{\alpha}^j$ is the step size, and $\tilde{S}^j$ is the vector representing the direction for the $j^{th}$ iteration and is given by [Appendix C, Eq. (C-37)]:

$$\tilde{S}^j = (1 - \beta) \frac{\tilde{\nabla} N^j}{\|\tilde{\nabla} N^j\|} + \beta \frac{\tilde{\nabla} N^{j-1}}{\|\tilde{\nabla} N^{j-1}\|} \quad , \qquad (23)$$

with $\beta = 0.35$. Here $\tilde{\nabla}$ represents the gradient vector. The step size $\tilde{\alpha}^j$ is set equal to

$$\alpha^j = K^j |\tilde{\phi}^j| \quad , \quad j \geq 0 \quad , \qquad (24)$$

where K is a parameter that is adjusted during any iteration to arrive at an optimal step size.

The approach for determining the power or frequency plans is best described with the help of the simplified conceptual flow chart diagram of the program STEEP shown in Figure 5. The program will stop if any one of the following three criteria is satisfied:

(1) The number of iterations equals (m), the maximum number of allowable iterations.

(2) $|N^{j+1} - N^j| \leq \varepsilon_1$.

(3) $[\sum_i (\phi_i^{j+1} - \phi_i^j)^2]^{1/2} \leq \varepsilon_2$.

The values for $\varepsilon_1$ and $\varepsilon_2$ are specified.

The program is started by specifying an initial starting value for $\tilde{\phi}$ and setting K = 1. During the $j^{th}$ iteration, the new value of $\tilde{\phi}$ is determined from Eq. (22), and a check is made to see whether $N^{j+1} < N^j$. If the condition is satisfied but the stopping criterion is not met, the value of K is multiplied by two in Loop 1 to increase the step size for the next iteration. If the condition $N^{j+1} < N^j$ is not satisfied, the step size in the previous iteration was too large. In this case, the

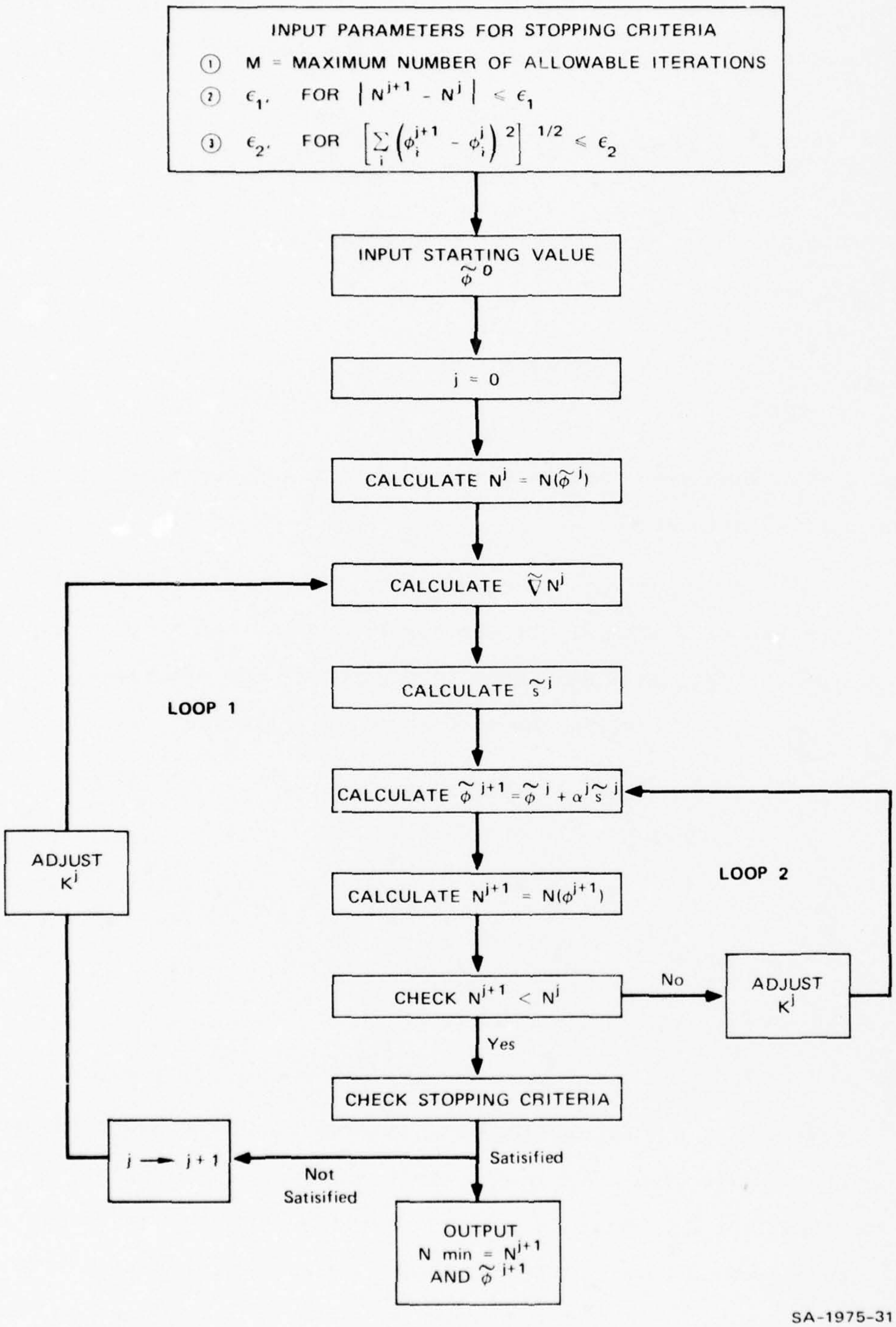


FIGURE 5 SIMPLIFIED FLOW CHART OF THE OPTIMIZATION PROGRAM STEEP

value of K is halved in Loop 2 to reduce the step size. This procedure is repeated until $N^{j+1} < N^{j}$ is satisfied. The value of K is thus adjusted continually in the two loops until any one of the criteria for stopping the program is met.

It should be mentioned that this approach for determining power and frequency plans provides the best plan for any selected ordering of the links in the repeater. Since some initial orderings of the links are expected to be better than others, it would be desirable to start with a good ordering before exercising the program STEEP for selecting power levels and frequencies. So far it has not been possible to develop general rules based on sound technical considerations for selecting the initial ordering of the links in the repeater when the FDMA operation involves a mix of users with differing characteristics.

### 3. Description of SYSCON

A computer program (SYSCON) has been developed to model a PSK/FDMA communication system and to optimize its performance when operating with a mix of user terminals through selection of power and frequency plans. The satellite transponder is modeled as a frequency translating and limiting repeater (Figure 1), whose limiting characteristic is described by a sum of two error functions [Eq. (1)]. The transmitters are characterized by their effective radiated powers (ERPs), RF frequencies, and data rates. Quadriphase PSK is assumed to be the data modulation for all the users. The receivers are characterized by their ratio of antenna gain to effective noise temperature (G/T). The computer program is described in detail in Appendix C.

At the start of the program it is essential to provide the initial ordering of the links in the satellite repeater, as well as the selection of the norm ($N_a$ or $N_b$) and the indication whether optimization

of the norm is desired. Also necessary is the information whether other possible combinations of the links are to be considered. If this is desired, the program is set up to consider all the possible link combinations and to determine the best power and frequency plan for each ordering. For a given number of links, it is therefore possible to obtain the power and frequency plans for all possible combinations. This allows selection of the ordering in the repeater that provides the minimum value of the norm.

At the end of each optimization (power or frequency), the probability of error for each channel, the best transmitter ERP/frequency assignments, and the value of the norm are printed.

### 4. Filtering Considerations

The effects of RF filtering in the ground terminals are not considered in SYSCON. The RF bandpass filter in the receiver eases the problem of initial synchronization by removing out-of-band signals and intermodulation products. When the receiver has been synchronized, the RF filter may not be critical, since the suppression of the out-of-band signals and intermodulation products is then accomplished by the low-pass or the integrate-and-dump filter following the product demodulator. The consideration of RF filtering at the transmitter is much more crucial, however, because it confines the RF spectrum of the transmitted signal to a finite bandwidth. Since this is not considered in SYSCON, the RF bandwidth of the signals is much wider, leading to large adjacent-channel interference between the links. Furthermore, frequency optimization primarily reduces the effects of cross products and has only little effect against adjacent channel interference. Consequently, the numerical results (Section V) obtained by exercising SYSCON do not show significant improvement from frequency optimization, because the reduction

in the effects of cross products is masked by the large adjacent-channel interference effects resulting from ignoring RF filtering in the transmitters. It is recommended that this be considered in future efforts in order to determine the improvement in FDMA performance from frequency optimization.

# V NUMERICAL RESULTS

The numerical results described in this section were generated with the values of the satellite parameters listed in Table 1. The values at $\gamma_1$ and $\gamma_2$ in Eq. (1) were determined (Appendix C, Section 3) by specifying that the limiter output power increases linearly with its input at low input power levels and has a prescribed power back-off at 0-dB normalized input power level.*

Table 1

PARAMETER VALUES USED IN NUMERICAL COMPUTATION

| Description | Parameter | Value | Unit |
|---|---|---|---|
| Satellite receiving antenna gain | $G_{sr}$ | 16.8 | dB |
| Satellite transmitting antenna gain | $G_{st}$ | 17.0 | dB |
| Satellite ERP | $P_{max}$ | 13.0 | dBW |
| Gain of the preamplifier | $G_a$ | 69.0 | dB |
| Gain of the TWT amplifier | $G_{twt}$ | 27.0 | dB |
| Satellite noise temperature | $T_s$ | 1690 | °K |
| Down-link midband frequency | $f_{mid}$ | 7.275 | GHz |
| Frequency offset for down-link transmission | $\Delta f$ | 0.725 | GHz |

*In the tables in this section, $\gamma_{10}$ and $\gamma_{20}$ refer to the corresponding values of $\gamma_1$ and $\gamma_2$, respectively, with respect to a maximum limiter output power of 0 dBW.

We considered first the case involving four links with two different data rates. All receivers were assumed to have the same figure of merit (G/T), equal to 44.5 dB/°K. The program was started by assigning initial power and frequency to the individual links proportional to their data rates. (This is achieved automatically by exercising the initialization plan described in Appendix C, Section 4.) The links carrying the lower data rate were assigned 74-dBW ERP, while the higher-data-rate links were given 80-dBW ERP. These values were determined such that the total power of all links at the limiter input would be 84 dBW. This value corresponds to a normalized value of 0 dB for the power of an equivalent sinusoidal signal at the limiter input. The limiter parameters $\gamma_1$ and $\gamma_2$ were determined so that the output power back-off equals 2.2 dB at 0-dB normalized input power level. The total repeater bandwidth was assumed to be 30 MHz, which provides a relatively high (81-percent) bandwidth utilization for FDMA operation.

The bit-error rate $P_{ei}$ for each channel before any optimization is shown in Table 2 under the heading "initial plan." Also given is the error rate ($P^{\star}_{ei}$) that would be obtained if only the contributions of the retransmitted up-link and down-link noise were to be considered at the output of each link. $P^{\star}_{ei}$ is given by

$$P^{\star}_{ei} = \frac{1}{2}\left[1 - \mathrm{erf}\left(\rho_i/\sqrt{2}\right)\right] \quad , \qquad (25)$$

where $\rho_i^2$ is the desired SNR for the $i^{th}$ channel. Since $\rho_i$ is a function of $A_i$ [Eq. (12)], it takes into account the power-sharing loss and signal suppression in the limiter. Equation (25) represents the error probability of an FDMA system operating with a limiting satellite repeater; it considers the effects of both up-link and down-link noise but neglects the effects of adjacent-channel interference and the cross products generated in the limiting repeater. A practical FDMA system will approach this

Table 2

OPTIMIZATION EXAMPLE 1

Repeater Bandwidth: 30 MHz
Limiter Parameters: $\gamma_{10} = 0.886$, $\gamma_{20} = 0.886$
Bandwidth Utilization: 81 percent

Order of Optimization: P,F,P,F
Initial Power Back-Off: 2.2 dB
Final Power Back-Off: 0.34 dB

Norm: $N_a$

| | | Initial Plan | | | | Final Plan | | | |
|---|---|---|---|---|---|---|---|---|---|
| $(G/T)_i$ (dB/°K) | $R_i$ (kbps) | $S_{ti}$ (dBW) | $f_i$ (GHz) | $P_{ei}$ | $P^*_{ei}$ | $S_{ti}$ (dBW) | $\Delta f_i$ (MHz) | $P_{ei}$ | $P^*_{ei}$ |
| 44.5 | 2424 | 74.0 | 7.986 | 1.749-3 | 5.042-4 | 82.8 | 0.000 | 1.346-5 | 6.997-7 |
| 44.5 | 9684 | 80.0 | 7.994 | 7.049-4 | 2.890-4 | 87.4 | -2.506 | 5.550-5 | 7.604-6 |
| 44.5 | 9684 | 80.0 | 8.006 | 1.749-4 | 2.890-4 | 87.6 | -1.041 | 3.780-5 | 4.332-6 |
| 44.5 | 2424 | 74.0 | 8.014 | 1.749-3 | 5.042-4 | 82.7 | 0.000 | 1.296-5 | 1.180-6 |
| Average | | | | 1.227-3 | 3.966-4 | | | 2.992-5 | 3.453-6 |

limit at large repeater bandwidths (low bandwidth utilization), because then it should be possible to place the signals at frequencies such that the cross products do not fall on the channels, and also there should be no adjacent-channel interference. In this case the error rate will be determined solely by the noise present in the system. Equation (25) serves as a convenient and meaningful reference point to compare the performance of a practical band-limited FDMA system.

The final power and frequency plans,* as well as the error rates obtained by exercising the optimization procedure in SYSCON, are shown in Table 2 under Final Plan. It is apparent that the improvement in error rate results primarily from the optimization on power and to a much lesser degree from optimization on frequency. The reason for this--as pointed out at the end of the previous section--is that frequency optimization greatly reduces the effects of cross products but only slightly reduces adjacent-channel interference. It was determined that the effect of adjacent-channel interference on error rate is, in general, much larger than the effect from the cross products. Thus, the reduction in the effect of cross products achieved by frequency optimization is masked by large adjacent-channel interference. If RF filtering for the individual links had been considered, the adjacent-channel interference would have been smaller, and it would then have been possible to see a more significant effect of frequency optimization.

As mentioned earlier, the program was initiated by assuming the sum of the ERPs of all the four links to be equal to 84 dBW. An equivalent sinusoid of this ERP will yield 0-dB normalized power at the limiter input.

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*In Table 2 and in all the tables following it, $\Delta f_i$ represents the frequency difference between the final and initial frequencies of the $i^{th}$ link. The final frequency is obtained by adding $\Delta f_i$ to the initial frequency.

The power back-off at this operating point equals 2.2 dB. Summing the total power of the four links in the final plan yields an equivalent power of 90.8 dBW for the equivalent sinusoid. This provides an increase of 6.8 dB in the normalized power level at the limiter input. For the assumed limiter characteristic, a 6.8-dB input power corresponds to an output power back-off of 0.4 dB. In Table 2 and the subsequent tables, both initial and final power back-offs are given.

Next, we considered the performance of all the other different combinations of these four links. Tables 3, 4, and 5 show the results. Again, it is evident that the effect of power optimization is much more significant than that of frequency optimization. From the four possible combinations, the best performance is achieved with the two higher data-rate links in the middle and the smaller data-rate links on each side. The next best combination is the two smaller data-rate links in the middle and the two larger data-rate links on the side. The combination with pairs of higher and lower data-rate links ranks third. Interspacing higher and lower data-rate links gives the worst performance of all the four possible combinations.

Table 6 shows results for the case involving four links with the same two data rates and initial frequency assignments as in Table 2, but we purposely assigned higher ERP to the lower data-rate links and smaller ERP to the higher data-rate links. We find that initially the error rate of the lower data-rate links is nearly perfect, while that of higher data-rate links is extremely poor. After optimization, we obtain good error rates for all four channels. As expected, the ERP of the lower data-rate links must be reduced, while that of the higher data-rate links must be increased, to obtain the best system error-rate performance. Note that the final power and frequency plans are reversed (by flipping the spectrum end for end) compared to the case shown in Table 2.

Table 3

OPTIMIZATION EXAMPLE 2

Repeater Bandwidth: 30 MHz

Limiter Parameters: $\gamma_{10} = 0.886$, $\gamma_{20} = 0.886$

Bandwidth Utilization: 81 percent

Order of Optimization: P,F,P,F

Initial Power Back-Off: 2.2 dB

Final Power Back-Off: 0.42 dB

Norm: $N_a$

| | | Initial Plan | | | | Final Plan | | | |
|---|---|---|---|---|---|---|---|---|---|
| $(G/T)_i$ (dB/°K) | $R_i$ (kbps) | $S_{ti}$ (dBW) | $f_i$ (GHz) | $P_{ei}$ | $P^*_{ei}$ | $S_{ti}$ (dBW) | $\Delta f_i$ (MHz) | $P_{ei}$ | $P^*_{ei}$ |
| 44.5 | 9684 | 80.0 | 7.990 | 5.001-4 | 2.890-4 | 86.3 | 0.000 | 8.419-5 | 1.018-5 |
| 44.5 | 2424 | 74.0 | 7.998 | 2.154-3 | 5.042-4 | 82.4 | -1.065 | 5.804-5 | 1.251-7 |
| 44.5 | 2424 | 74.0 | 8.002 | 2.154-3 | 5.042-4 | 82.0 | -1.313 | 4.069-5 | 4.094-7 |
| 44.5 | 9684 | 80.0 | 8.010 | 5.001-4 | 2.890-4 | 86.0 | 0.000 | 8.194-5 | 2.078-5 |
| Average | | | | 1.327-3 | 3.996-4 | | | 6.622-5 | 7.873-6 |

Table 4

OPTIMIZATION EXAMPLE 3

Repeater Bandwidth: 30 MHz
Limiter Parameters: $\gamma_{10} = 0.886$, $\gamma_{20} = 0.886$
Bandwidth Utilization: 81 percent

Order of Optimization: P,F,P,F
Initial Power Back-Off: 2.2 dB
Final Power Back-Off: 0.45 dB

Norm: $N_a$

| | | Initial Plan | | | | Final Plan | | | |
|---|---|---|---|---|---|---|---|---|---|
| $(G/T)_i$ (dB/°K) | $R_i$ (kbps) | $S_{ti}$ (dBW) | $f_i$ (GHz) | $P_{ei}$ | $P^{\star}_{ei}$ | $S_{ti}$ (dBW) | $\Delta f_i$ (MHz) | $P_{ei}$ | $P^{\star}_{ei}$ |
| 44.5 | 9684 | 80.0 | 7.988 | 5.363-4 | 2.890-4 | 85.8 | 0.000 | 6.451-5 | 1.429-5 |
| 44.5 | 9684 | 80.0 | 8.001 | 7.524-4 | 2.890-4 | 85.9 | 1.083 | 1.370-4 | 1.222-5 |
| 44.5 | 2424 | 74.0 | 8.009 | 1.989-3 | 5.042-4 | 81.7 | 0.560 | 6.321-5 | 3.785-7 |
| 44.5 | 2424 | 74.0 | 8.012 | 1.298-3 | 5.042-4 | 81.3 | 0.000 | 3.056-5 | 1.447-6 |
| Average | | | | 1.144-3 | 3.966-4 | | | 7.382-5 | 7.084-6 |

Table 5

OPTIMIZATION EXAMPLE 4

Repeater Bandwidth: 30 MHz
Limiter Parameters: $\gamma_{10} = 0.886$, $\gamma_{20} = 0.886$
Bandwidth Utilization: 81 percent

Order of Optimization: P,F,P,F
Initial Power Back-Off: 2.2 dB
Final Power Back-Off: 0.5 dB

Norm: $N_a$

| | | Initial Plan | | | | Final Plan | | | |
|---|---|---|---|---|---|---|---|---|---|
| $(G/T)_i$ (dB/°K) | $R_i$ (kbps) | $S_{ti}$ (dBW) | $f_i$ (GHz) | $P_{ei}$ | $P^*_{ei}$ | $S_{ti}$ (dBW) | $\Delta f_i$ (MHz) | $P_{ei}$ | $P^*_{ei}$ |
| 44.5 | 9684 | 80.0 | 7.988 | 5.383-4 | 2.890-4 | 85.4 | 0.000 | 1.536-4 | 2.606-5 |
| 44.5 | 2424 | 74.0 | 7.996 | 2.973-3 | 5.042-4 | 81.6 | 0.278 | 1.036-4 | 2.738-7 |
| 44.5 | 9684 | 80.0 | 8.004 | 6.757-4 | 2.890-4 | 85.8 | -1.409 | 8.470-5 | 8.147-6 |
| 44.5 | 2424 | 74.0 | 8.012 | 1.495-4 | 5.042-4 | 80.9 | 0.000 | 1.779-5 | 2.194-6 |
| Average | | | | 1.420-3 | 3.966-4 | | | 8.990-5 | 9.168-6 |

Table 6

OPTIMIZATION EXAMPLE 5

Repeater Bandwidth: 30 MHz
Limiter Parameters: $\gamma_{10} = 0.886$, $\gamma_{20} \approx 0.886$
Bandwidth Utilization: 81 percent

Order of Optimization: P,F,P,F
Initial Power Back-Off: 2.2 dB
Final Power Back-Off: 0.38 dB

Norm: $N_a$

| | | Initial Plan | | | | Final Plan | | | |
|---|---|---|---|---|---|---|---|---|---|
| $(G/T)_i$ (dB/°K) | $R_i$ (kbps) | $S_{ti}$ (dBW) | $f_i$ (GHz) | $P_{ei}$ | $P^*_{ei}$ | $S_{ti}$ (dBW) | $\Delta f_i$ (MHz) | $P_{ei}$ | $P^*_{ei}$ |
| 44.5 | 2424 | 80.0 | 7.986 | 1.441-11 | 3.217-12 | 82.5 | 0.000 | 1.312-5 | 7.803-7 |
| 44.5 | 9684 | 74.0 | 7.994 | 7.316-2 | 4.988-2 | 87.1 | 0.971 | 5.975-5 | 7.358-6 |
| 44.5 | 9684 | 74.0 | 8.006 | 7.316-2 | 4.988-2 | 87.2 | 2.494 | 5.004-5 | 6.783-6 |
| 44.5 | 2424 | 80.0 | 8.014 | 1.441-11 | 3.217-12 | 82.6 | 0.000 | 1.370-5 | 6.706-7 |
| Average | | | | 3.657-2 | 2.494-2 | | | 3.415-5 | 3.900-6 |

Tables 7 to 9 show the results for the case involving six links with three different data rates and receiver figures of merit. The initial power and frequency ordering was determined from the initialization plan. The limiter characteristic shown in Figure 4 was used for this example. The initial power plan corresponds to a 4-dB back-off. Table 7 shows the results of optimization with a repeater bandwidth of 40 MHz, corresponding to a bandwidth utilization of 66 percent. Tables 8 and 9 give the difference in the system performance when the bandwidth is increased to 80 MHz and 160 MHz, corresponding to reductions in bandwidth utilization of 33 percent and 16.5 percent. It is interesting to compare the average error rate of the band-limited FDMA system with that of an "ideal" FDMA system in the three tables. It is clearly evident that, as the repeater bandwidth is increased, the performance of a practical FDMA system rapidly approaches that of an ideal system, as expected. Because of increased available bandwidth, it is now possible to select frequency assignments so that the effects of cross products and adjacent channel interference are greatly reduced.

Table 10 shows the results of a six-signal case, for which the initial conditions were the same as the case considered in Table 7; but for optimization the norm $N_b$ was used instead of $N_a$. Norm $N_b$ considers the sum of the error probabilities weighted by their data rates. Since the two center channels carry higher data rates, they were weighted more heavily than the other channels. Comparison of the final plans in Tables 7 and 10 shows that, because of this weighting, the error rates of the two center channels have improved at the expense of the other channels. Also, the average (arithmetic mean) error rate of the system has degraded. The frequency plan, however, has remained unchanged.

Table 7

OPTIMIZATION EXAMPLE 6

Repeater Bandwidth: 40 MHz

Limiter Parameters: $\gamma_{10} = 0.541$, $\gamma_{20} = 2.445$

Bandwidth Utilization: 66 percent

Order of Optimization: P,F,P,F

Initial Power Back-Off: 4 dB

Final Power Back-Off: 0.9 dB

Norm: $N_a$

| | | Initial Plan | | | | Final Plan | | | |
|---|---|---|---|---|---|---|---|---|---|
| $(G/T)_i$ (dB/°K) | $R_i$ (kbps) | $S_{ti}$ (dBW) | $f_i$ (GHz) | $P_{ei}$ | $P^*_{ei}$ | $S_{ti}$ (dBW) | $\Delta f_i$ (MHz) | $P_{ei}$ | $P^*_{ei}$ |
| 35.5 | 1200 | 76.2 | 7.981 | 1.224-2 | 1.119-2 | 86.0 | 0.000 | 2.860-3 | 2.223-3 |
| 38.5 | 2400 | 76.2 | 7.983 | 1.320-2 | 1.130-2 | 86.2 | 0.066 | 2.760-3 | 1.747-3 |
| 44.5 | 9600 | 76.2 | 7.993 | 1.700-2 | 1.199-2 | 86.6 | 0.385 | 4.161-3 | 1.163-3 |
| 44.5 | 9600 | 76.2 | 8.007 | 1.700-2 | 1.199-2 | 87.0 | 0.562 | 2.580-3 | 6.384-4 |
| 38.5 | 2400 | 76.2 | 8.017 | 1.320-2 | 1.130-2 | 86.2 | 0.462 | 2.743-3 | 1.781-3 |
| 35.5 | 1200 | 76.2 | 8.019 | 1.224-2 | 1.119-2 | 85.8 | 0.000 | 3.491-3 | 2.730-3 |
| Average | | | | 1.414-2 | 1.149-2 | | | 3.098-3 | 1.713-3 |

Table 8

OPTIMIZATION EXAMPLE 7

Repeater Bandwidth: 80 MHz
Limiter Parameters: $\gamma_{10} = 0.541$, $\gamma_{20} = 2.445$
Bandwidth Utilization: 33 percent

Order of Optimization: P,F,P,F
Initial Power Back-Off: 4 dB
Final Power Back-Off: 0.6 dB

Norm: $N_a$

| | | Initial Plan | | | | Final Plan | | | |
|---|---|---|---|---|---|---|---|---|---|
| $(G/T)_i$ (dB/°K) | $R_i$ (kbps) | $S_{ti}$ (dBW) | $f_i$ (GHz) | $P_{ei}$ | $P^*_{ei}$ | $S_{ti}$ (dBW) | $\Delta f_i$ (MHz) | $P_{ei}$ | $P^*_{ei}$ |
| 35.5 | 1200 | 76.2 | 7.961 | 1.179-2 | 1.149-2 | 87.7 | 0.000 | 1.671-3 | 1.542-3 |
| 38.5 | 2400 | 76.2 | 7.966 | 1.224-2 | 1.160-2 | 87.8 | -0.187 | 1.714-3 | 1.425-3 |
| 44.5 | 9600 | 76.2 | 7.985 | 1.342-2 | 1.231-2 | 88.0 | -0.903 | 1.887-3 | 1.146-3 |
| 44.5 | 9600 | 76.2 | 8.015 | 1.342-2 | 1.231-2 | 88.1 | -1.481 | 1.850-3 | 1.111-3 |
| 38.5 | 2400 | 76.2 | 8.034 | 1.224-2 | 1.160-2 | 87.8 | -1.451 | 1.699-3 | 1.521-3 |
| 35.5 | 1200 | 76.2 | 8.039 | 1.179-2 | 1.149-2 | 87.8 | 0.000 | 1.699-3 | 1.582-3 |
| Average | | | | 1.248-2 | 1.180-2 | | | 1.748-3 | 1.387-3 |

Table 9

OPTIMIZATION EXAMPLE 8

Repeater Bandwidth: 160 MHz
Limiter Parameters: $\gamma_{10} = 0.541$, $\gamma_{20} = 2.445$
Bandwidth Utilization: 16.5 percent

Order of Optimization: P,F,P,F
Initial Power Back-Off: 4 dB
Final Power Back-Off: 0.5 dB

Norm: $N_a$

| | | Initial Plan | | | | Final Plan | | | |
|---|---|---|---|---|---|---|---|---|---|
| $(G/T)_i$ (dB/°K) | $R_i$ (kbps) | $S_{ti}$ (dBW) | $f_i$ (GHz) | $P_{ei}$ | $P^*_{ei}$ | $S_{ti}$ (dBW) | $\Delta f_i$ (MHz) | $P_{ei}$ | $P^*_{ei}$ |
| 35.5 | 1200 | 76.2 | 7.921 | 1.226-2 | 1.210-2 | 87.7 | 0.000 | 1.771-3 | 1.731-3 |
| 38.5 | 2400 | 76.2 | 7.932 | 1.254-2 | 1.222-2 | 88.2 | -0.260 | 1.271-3 | 1.185-3 |
| 44.5 | 9600 | 76.2 | 7.970 | 1.355-2 | 1.294-2 | 88.4 | -1.582 | 1.178-3 | 9.235-4 |
| 44.5 | 9600 | 76.2 | 8.030 | 1.355-2 | 1.294-2 | 88.3 | -2.275 | 1.547-3 | 1.234-3 |
| 38.5 | 2400 | 76.2 | 8.068 | 1.254-2 | 1.222-2 | 88.1 | -1.816 | 1.635-3 | 1.547-3 |
| 35.5 | 1200 | 76.2 | 8.079 | 1.226-2 | 1.210-2 | 88.0 | 0.000 | 1.828-3 | 1.796-3 |
| Average | | | | 1.278-2 | 1.242-2 | | | 1.538-3 | 1.402-3 |

Table 10

OPTIMIZATION EXAMPLE 9

Repeater Bandwidth: 40 MHz
Limiter Parameters: $\gamma_{10} = 0.541$, $\gamma_{20} = 2.445$
Bandwidth Utilization: 66 percent

Order of Optimization: P,F,P,F
Initial Power Back-Off: 4 dB
Final Power Back-Off: 1 dB

Norm: $N_b$

| | | Initial Plan | | | | Final Plan | | | |
|---|---|---|---|---|---|---|---|---|---|
| $(G/T)_i$ (dB/°K) | $R_i$ (kbps) | $S_{ti}$ (dBW) | $f_i$ (GHz) | $P_{ei}$ | $P^*_{ei}$ | $S_{ti}$ (dBW) | $\Delta f_i$ (MHz) | $P_{ei}$ | $P^*_{ei}$ |
| 35.5 | 1200 | 76.2 | 7.981 | 1.224-2 | 1.119-2 | 84.4 | 0.000 | 8.844-3 | 7.251-3 |
| 38.5 | 2400 | 76.2 | 7.983 | 1.320-2 | 1.130-2 | 85.5 | 0.001 | 3.915-3 | 2.271-3 |
| 44.5 | 9600 | 76.2 | 7.993 | 1.700-2 | 1.199-2 | 87.3 | 0.001 | 1.156-3 | 2.368-4 |
| 44.5 | 9600 | 76.2 | 8.007 | 1.700-2 | 1.199-2 | 87.4 | 0.001 | 9.973-4 | 1.989-4 |
| 38.5 | 2400 | 76.2 | 8.017 | 1.320-2 | 1.130-2 | 85.5 | -0.002 | 3.939-3 | 2.736-3 |
| 35.5 | 1200 | 76.2 | 8.019 | 1.224-2 | 1.119-2 | 84.4 | 0.000 | 8.915-3 | 7.305-3 |
| Average | | | | 1.414-2 | 1.143-2 | | | 4.627-3 | 3.333-3 |

Finally, we considered a case involving a mix of ten user terminals with differing up-link ERPs, data rates, and receiving capabilities. The results with initial and final power and frequency plans after optimization are shown in Table 11.

Table 11

OPTIMIZATION EXAMPLE 10

Repeater Bandwidth: 185 MHz
Limiter Parameters: $\gamma_{10} = 0.886$, $\gamma_{20} = 0.886$
Bandwidth Utilization: 12.7 percent

Order of Optimization: P,F,P,F
Initial Power Back-Off: 0.5 dB
Final Power Back-Off: 0.2 dB

Norm: $N_a$

| | | Initial Plan | | | | Final Plan | | | |
|---|---|---|---|---|---|---|---|---|---|
| $(G/T)_i$ (dB/°K) | $R_i$ (kbps) | $S_{ti}$ (dBW) | $f_i$ (GHz) | $P_{ei}$ | $P^*_{ei}$ | $S_{ti}$ (dBW) | $\Delta f_i$ (MHz) | $P_{ei}$ | $P^*_{ei}$ |
| 49.2 | 64.8 | 58.1 | 7.908 | 4.028-4 | 2.718-4 | 64.10 | 0.000 | 6.557-7 | 1.508-7 |
| 44.2 | 648 | 73.1 | 7.910 | 2.659-4 | 2.190-4 | 77.50 | 0.006 | 1.101-5 | 8.383-6 |
| 39.2 | 648 | 78.1 | 7.915 | 1.964-4 | 1.847-4 | 81.80 | 0.120 | 2.999-5 | 2.796-5 |
| 39.2 | 1940 | 82.9 | 7.926 | 1.231-4 | 1.164-4 | 85.85 | 0.920 | 7.177-5 | 6.839-5 |
| 39.2 | 3240 | 85.1 | 7.946 | 6.913-5 | 6.707-5 | 87.39 | 5.230 | 1.553-4 | 1.512-4 |
| 49.2 | 12970 | 81.1 | 8.010 | 3.195-4 | 1.978-4 | 84.60 | 9.101 | 5.686-5 | 3.896-5 |
| 29.2 | 97.30 | 79.9 | 8.061 | 1.557-4 | 1.555-4 | 83.30 | 2.821 | 4.224-5 | 4.210-5 |
| 44.2 | 3240 | 80.1 | 8.075 | 1.954-4 | 1.729-4 | 83.60 | 2.432 | 4.430-5 | 4.076-5 |
| 39.2 | 648 | 78.1 | 8.090 | 1.918-4 | 1.847-4 | 81.80 | 1.621 | 2.998-5 | 2.883-5 |
| 39.2 | 6.48 | 58.1 | 8.092 | 2.365-4 | 2.178-4 | 63.60 | 0.000 | 5.988-7 | 5.224-7 |
| Average | | | | 2.156-4 | 1.787-4 | | | 4.427-5 | 4.072-5 |

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13. ABSTRACT

This report presents the results of research performed to investigate the application of the frequency-division multiple-access (FDMA) technique for providing a mix of user terminals of differing characteristics--such as data rates, transmitter powers, and receiver sensitivities--with simultaneous access capability to a limiting satellite repeater. A computer program (SYSCON) has been developed to model a PSK/FDMA satellite communication system and to optimize its performance in operation with a mix of user terminals, through selection of power and frequency plans. This capability is achieved through optimization of a norm defined as the weighted sum of the link error rates and representing the figure of merit or a measure of the system's communications performance with respect to both the power and the frequency of the links. The method of steepest descent is used for determining both power and frequency plans. It was found that through power and frequency control the limiting satellite repeater can be operated in the saturation region at substantially higher power levels (1-dB back-off) than is customary in practice.

To obtain the expression for the error rate at the input of the FDMA links, it was necessary to derive general analytic expressions for the limiter output signal, the intermodulation, and the noise components, when n signals are transmitted simultaneously through the satellite repeater. The expression for the bit error rate was then derived by assuming digital quadriphase modulation of the FDMA carriers and taking into consideration the presence of other FDMA carriers causing adjacent-channel interference, the intermodulation products generated in the limiter, and retransmitted satellite repeater noise, as well as receiver noise.

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| KEY WORDS | LINK A | | LINK B | | LINK C | |
|---|---|---|---|---|---|---|
| | ROLE | WT | ROLE | WT | ROLE | WT |
| Limiters, frequency-division multiple access, intermodulation products, error rate, steepest descent, time-division multiple access, spread-spectrum multiple access, power control, frequency control | | | | | | |

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13. Abstract (Concluded)

A general comparison of the three major multiple-access alternatives--FDMA, TDMA, and SSMA--with respect to selected performance criteria that are particularly important for the military environment and for operation with a mix of users indicated that FDMA performs much better than past analyses had shown. With the same satellite power and RF bandwidth, FDMA was found to offer nearly as much satellite throughput as TDMA and considerably more than SSMA.

The report consists of three volumes. Volume One provides a summary of the study which includes a description of the analysis approach, documentation of the pertinent equations, numerical results, and conclusions. Detailed analysis and investigation of the problem areas, as well as derivations of analytical expressions, are contained in the appendices in Volume Two. Volume Three provides a technical assessment of the suitability of PSK/FDMA for operation with a mix of users and compares its performance with other multiple-access alternatives, in particular, with that of time-division multiple access (TDMA) and spread-spectrum multiple access (SSMA).